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Application No.   S2002/0985

Date of Filing   20th December 2002

Applicant   The Provost fellows and scholars of the College of the Holy and Undivided Trinity of Queen Elizabeth near Dublin of College Green, Dublin 2, Ireland

Dated this 28TH day of February 2003.

PRIORITY DOCUMENT

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FORM NO. 1

REQUEST FOR THE GRANT OF A PATENT

PATENTS ACT 1992

The Applicant(s) named herein hereby request(s)
    [   ] the grant of a patent under Part II of the Act
    [X] the grant of a short-term patent under Part III of the Act
on the basis of the information furnished hereunder.

1. Applicant(s)

THE PROVOST FELLOWS AND SCHOLARS OF THE COLLEGE OF THE HOLY AND
UNDIVIDED TRINITY OF QUEEN ELIZABETH NEAR DUBLIN
of College Green
Dublin 2
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2. Title of Invention
Device and method for inspection of high frequency and microwave
hybird circuits and printed circuit boards

3. Declaration of Priority on basis of previously filed application(s) for same invention (Sections 25 & 26)

Previous Filing Date	Country in or for which filed	Filing No.

4. Identification of Inventor(s)

    Name(s) and address(es) of person(s) believed
    by the Applicant(s) to be the inventor(s)
      To Follow

5. **Statement of right to be granted a patent (Section 17(2) (b))**

To follow

6. **Items accompanying this Request**

(i)   [X] prescribed filing fee (Euro 60.00)
(ii)   [ ] specification containing a description and claims
         [X] specification containing a description only
         [X] Drawings referred to in description or claims
(iii)   [ ] An abstract
(iv)   [ ] Copy of previous application(s) whose priority is
              claimed
(v)   [ ] Translation of previous application whose
              priority is claimed
(vi)   [X] Authorisation of Agent (this may be given at 8
              below if this Request is signed by the Applicant(s))

7. **Divisional Application(s)**

The following information is applicable to the present application which is made under Section 24 -

Earlier Application No.
          Filing Date:

8. **Agent**

The following is authorised to act as agent in all proceedings connected with the obtaining of a patent to which this request relates and in relation to any patent granted -
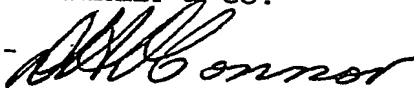
Name & Address

Cruickshank & Co. at their address recorded for the time being in the Register of Patent Agents is hereby appointed Agents and address for service, presently 1 Holles Street, Dublin 2.

9. **Address for service (if different from that at 8)**

Signed Cruickshank & Co.

By:-   Executive.

Agents for the Applicant

Date December 20, 2002.

Device and method for inspection of high frequency and microwave hybrid circuits and printed circuit boards.

Introduction

Printed Circuit Boards (PCB) and hybrid circuit assemblies need to be tested during both the development and production phases. During the development phase the tests should allow the establishment of errors in the circuit design, confirm the correctness of the choice of elements and the optimisation of the circuit layout. To answer these questions one needs to collect rather detailed information on currents or voltages in the circuit during its operation, their phase and spectral characteristics. During the production phase, one often needs to perform relatively simple measurements at a number of points on a circuit and compare these with readings for a correctly functioning one to enable quality control and in some cases component adjustment procedure. Therefore, requirements to the test technology are rather different during the development and the production phase. During the development phase, in-depth information is required on a relatively small number of units. The units could be an electronic circuit, PCB, hybrid circuit or any other similar circuit hereinafter called Device Under Test (DUT). During the production phase one does not have to obtain in-depth information but the measurements have to be performed rapidly so that a large number of DUTs could pass through the quality control procedure.

At present the standard method of testing employed measures the voltage signal at a number of test points by applying probes to these. There are technologies enabling the establishment of contacts between the DUT and the test rig through a large number of spring-loaded probes simultaneously. These are called the "Bed of Nails" test technologies. Various fixtures have been developed for the "Bed of Nail" and related technologies. The prior art is represented by for example in US Pat. Nos 4,017,793- Haines; US 4,056,733-Sullivan; US 4,061,969- Dean; US 4,115,735- Stanford; 4,164,704- Kato et al; US 4,209,745 Hines; US 4,321,533- Matrone; US 4,322,682 Schadwill; US 5,216,358. Such technologies work reliably at a low frequency range, generally up to 100 MHz. For testing of unpowered PCB resistance is usually measured between various tracks of the board. In some cases capacitance is measured between the tracks and the ground layer (e.g. US 4,583,042- Reimer). In the case of powered PCB, voltages are usually measured at the contact probes. Generally a similar approach is taken for testing high frequency and microwave circuits. The high frequency probes are more complex and difficult to use. The high frequency contact test technologies are described in US 4,697,143; US 4593243; US 5,565,788. In the case of high frequency and microwave DUTs such measurements are much more complex and cumbersome for a number of reasons:

1. Measurements require the setting up of special contact pads on the DUT thus imposing an additional design requirement. In the case of low frequency measurements, almost any pin of a device/printed circuit board can serve as a contact pad as it is. In the case of high frequency and microwave measurements, contact pads are much larger in size and have to be specially placed on the DUT just for the purpose of such testing.
2. The high frequency and microwave probes are much more complex than the low-frequency ones. To place the probe on the DUT correctly, a single probe usually has to make contact

with the DUT at two or three points simultaneously and not just at one point as with a low frequency probe. Therefore surface roughness of the DUT or contamination can play a detrimental role.

3. The high frequency probes are much more complex, fragile and less durable than the low-frequency ones.

4. Impedance of the probe has to be matched at the test point not to affect the circuit's operation. This imposes an additional design restriction on the probe and also on the DUT: meaning that either the DUT has to be designed is such a way that all the test points have identical impedance or that several different probes have to be used to test the DUT at several points.

5. The high frequency/microwave probe is much more likely to affect the performance of the DUT than probes taking low frequency signals.

The objective of the present invention is to address these shortcomings of the test technology at high and microwave frequencies. Primarily we aim at the frequency range from 50 MHz to 50 GHz. The intention is based on a radical departure from the key requirement of the current testing technology: establishment of a low-resistance contact between the high frequency/microwave probe and the DUT. In the conventional test technology if the resistance between the probe and the DUT increases, the results of the measurements become meaningless. In this invention we aim to separate the probe and the DUT by a distance gap, leading to very week coupling between them. The measurements are based on an antenna that detects the non-radiative electric and magnetic field emitted by the DUT in the near field region.

There are technologies where a PCB (generally an unpowered board) is placed in external electromagnetic field. The board perturbs the field. The pattern of field distortion contains information about the defects in tracks of the board. This technology was developed for finding faults in unpopulated or inactive populated PCBs. The field perturbation is measured by an array of electromagnetic sensors. An example of such technology is described e.g. in US Pat. 5,424,633- Soiferman. A relatively similar technology is described in US Pat. 5,006,788- Goulette et al.

There are inventions where the sample is scanned with respect to the probe in the near-field proximity of the probe. For example, the US patent specification US 5,781018 (Davidov et al) teaches a method of characterising properties of materials such as dielectric constant or local resistivity. In this technique the microwave signal is coupled through a wave-guide probe towards the sample. The signal is reflected from the sample back into the wave-guide. Two signals at two orthogonal polarisations are compared. A similar technique is described in US 6,100,703. Although these techniques could be beneficial for the testing of passive materials such as silicon wafers, they cannot be directly applied for testing active DUTs. Besides, although these techniques are capable of detecting relatively small features such as long scratches on a flat conducting surface, they cannot deliver resolution below 100 μm that is required for testing of PCB and hybrid circuits. From the data presented in the specification, one could estimate that the resolution is in the range of several millimetres. This is understandable as the technique utilises wave-guides that would be of a rather large size at the frequency of interest to us. It is clear that preferably, PCB test technology should not be based on a particular resonance frequency determined by the size of the probe.

There are numerous other inventions related to the same aspect: characterisation of material properties of the sample (such as local conductivity or dielectric constant) at high frequency by

bringing an open ended probe in close proximity to the material. Those techniques are presented in e.g patent specification US 5,900,618 (S.M. Anagle et al.) and publications [C. P. Vlahacos, R. C. Black, S. M. Anlage, A. Amar, F. C. Wellstood, Appl. Phys. Lett. 69 (1996), p.3272], [D. E. Steinhauer, C. P. Vlahacos, S. K. Dutta, F. C. Wellstood, S. M. Anlage, Appl. Phys. Lett. 71 (1996), p. 1736], [A. Kramer, F. Keilmann, B. Knoll, R. Gickenberger, Micron, Vol 27 (1997), p.413]. Electromagnetic field is coupled into the sample either from the probe or from an external source. The energy, reflected back from the sample into the probe or transmitted through the sample, is measured. A similar technique is described in US Pat. 6,173,604- Xiang et. al. The microwave energy is coupled into probe placed in proximity of the sample. The energy reflected from the sample back into the probe contains information about the sample properties such as dielectric constant. To improve the sensitive of the technique, the probe is placed in quarter wavelength cavity resonator. These techniques are based on monitoring relatively strong capacitive coupling between the probe and the sample. It is exactly for this reason that they can not be directly used for the testing of microwave devices and PCB, as preferably one tries to reduce the coupling of the probe to the DUT, as such coupling affects the performance of the DUT.

Due to good sensitivity and well-defined electric properties, short cylindrical coaxial antennas are commonly used for the acquisition of microwave electric intensities in a near-field region [J. S.Dahele, A. L. Cullen, IEEE Trans. Mic. Theory Tech. 28 (7), p.752 (1980); J. Gao, A. Lauder, Q. Ren, Wolf I., IEEE Trans. Mic. Theory Tech 46 (1998), p.1694, patent specification US 5,900,618 (S.M. Anagle et al.). Also known as short monopoles, they consist of a central conductor that protrudes for a defined length from the shielding. Because of the axial symmetry, such antennas are sensitive to the component of the electric field intensity parallel to that axis. The external field is commonly assumed to be homogeneous thus resulting in a single sensitivity coefficient, that is the ratio between the signal level induced in the antenna and the field intensity. The length of the protruding conductor must not exceed the desired spatial resolution. The resolution does not just depend on this length but also on the dimensions of the shielding as surface currents in the shielding induce a secondary field and change the input signal. When the field is highly localised around the apex of the protruding conductor, images with spatial resolution somewhat better than the length of the conductor and the dimensions of the shielding can be obtained. On the other hand those images lack good quantitative characterization as the antenna's signal level depends on a particular distribution of the field that can no longer be considered to be homogeneous. Additionally, when the antenna length is chosen to be comparable or shorter than the shield diameter, the presence of the shielding close to the circuit may cause redistribution of the charges in the circuit and distortion of the primary induced field.

For many DUTs it is beneficial to be able to perform the measurements with the spatial resolution of 100 micrometres or better as the width of the track on a PCB is in that range. It appears that only by decreasing the antenna dimensions along with the coaxial shielding its spatial resolution capability can then be improved. Therefore, one expects that in order to make the antenna a suitable basis of the test technology, it has to be made with dimensions smaller than 50 micrometres. There are, however, numerous complications in making such an antenna. It is difficult to manufacture reproducibly the antenna alone with such small dimensions, especially manufacturing the coaxial line with a small diameter and forming a short protruding central conductor. It is also difficult to establish a reliable and well-defined shield. Such a shield strongly influences the properties of any antenna. The second reason making it unattractive to reduce the antenna size into the micrometer range is that with decreasing the antenna size, its impedance values move away from the common

values of the microwave and radio frequency amplifiers. As a result, it is more difficult to couple the signal from the antenna into the preamplifier.

The objective of the present invention is to overcome the resolution limit determined by the antenna's dimensions and and increase its resolution capability without the need for further miniaturization of the antenna.

The spatial resolution also depends on the gap separating the antennas and the DUT. For large separation, the resolution gets worse. Also the signal detected by the antenna depends on this separation and again the larger the separation, the smaller the signal detected. It is, therefore, intuitively attractive to reduce the separation to a value as small as possible. This is not, however, the best course of action for the following reason: with a small separation the antenna starts influencing the DUT, mainly through the capacitive coupling between them. The situation with a ·small separation between the antenna and the surface of the DUT is effectively equivalent to a capacitor between the DUT and the antenna at the point of test. The smaller the separation, the greater the capacitor's value. This capacitance depends not only on the separation between the probe and the surface, but also on the dielectric properties of the material underneath the probe. The optimal separation between the probe and the sample must satisfy both criteria, enabling to achieve high resolution and low coupling between the probe and the sample. The separation should not exceed the desired resolution but it should be greater than the diameter of the protruding end of the antenna. The range of some 5-50 μm is practical for protruding wire conductors with diameters of similar values. The situation is further complicated by the fact most DUTs are not flat and contain complex topographical features such as electronic elements, wires, air bridges etc. Therefore, the intrinsic part of the invention is the method and means for controlling the separation between the end of the antenna and the DUT and maintaining this distance in that range with small error.

There is extensive state-of-the-art in the area of measuring topography of sample using techniques of scanning atomic force microscopy and scanning shear force microscopy described in patent literature and open publications – US 5,412,980, [P. C. Yang, Y. Chen, M. Vanez-Iravani, J. Appl. Phys. 71 (1992), p.2499], [R. Tolledo-Crow, P. C., Yang, Y. Chen, M. Varez-Iravani, Appl. Phys. Lett. 60 (1992), p.2957], [Y. Martin, C. C. Williams, H. K. Wickramasinghe, J. Appl. Phys. 62 (1997), p.4723]. Unfortunately many of these inventions are not suitable for testing of PCBs with microwave probes. Most of these topography scanning techniques utilise very light topography probes, such as silicon cantilevers, usually fabricated by microfabrication processes. In some cases combination of the topographic and field probe is done with the help of advances in microfabrication technologies whereby the functional electric or magnetic field probe is fabricated directly at the AFM cantilever. The topography probes are mostly based on mechanical resonance of the probe and changes in resonance conditions caused by the sample proximity. As these probes operate very close to the sample (in the range of 1-50 nm), they may be also sensitive to the electric or magnetic interaction between the sample and the probe, such as disclosed in US Pat. No 5,936,237 (Van der Weide, D. Warren). The probes can be withdrawn out of the sample for measurements of various probe-sample interactions at higher distances, as in system described in the US specification US 5,418,363 (Elings et al.) and references therein. The small mass of the probe is crucial to assure good sensitivity in scanning atomic force microscopy and shear force microscopy. Some of the probes combine the tip probe with integral small electric or magnetic field antenna. Van der Weide, D. Warren, (US 5,936,237 patent specification) teaches the combination of the electromagnetic probe with the probe of Atomic Force Microscope (AFM) in a single multipurpose probe. The electromagnetic measurements are based on coupling between the probe and the sample and are suitable for testing materials at the frequencies up to above THz. Although we do not dispute that it may be possible to fabricate micrometer size probes capable of providing

high resolution ░░░░░░░░░, we do know from our experience that the use of such miniature probes for PCB testing in the microwave frequency range (50 MHz to 50 GHz) with any kind of satisfactory performance is virtually impossible. This is due to the fundamental relation between the probe size and the sensitivity of the probe for the frequencies of interest and also practical requirements put on the DUT flatness and size. These reasons limit possible use of such existing technologies to the inspection of semiconductor wafers and the study of materials properties of relatively flat samples.

Detailed description of the invention

The invention will be more clearly understood from the description of the following figures:

The embodiment utilising a coaxial antenna is shown in Fig. 1a. Disclosed methods allow measurement of the fields with high resolution and low perturbation of measured field without the need of reducing the length of the antenna's coaxial protruding conductor and reducing the diameter of the shield. On the contrary, the protruding conductor is relatively long. It is crucially important that unlike in many state of the art descriptions, this protruding conductor substantially exceeds both the diameter D (typically $D = 0.1$-0.5 mm) of the coaxial shield and the size w of measured signal lines of the DUT, typically

$$l > 3D, 3w \qquad (1)$$

whichever (D or w) is greater. Such a configuration allows low perturbation of the signals in DUT as the shielding body is relatively distant from the measured sample. The reason for this is that if the length of the protruding conductor is too short and the distance between the shielding and the signal line compares or is smaller than their dimensions, the shield starts affecting the signals within DUT's and changes its performance. Usual lengths l and the diameters d of the central protruding conductor are $l = 0.2$-3 mm respectively $d = 5$-$50\mu m$ for the transmission lines within the range of $30\mu m - 0.5mm$, but may be bigger for larger structures. On the other hand, as we have described above, one should expect that the resolution should be comparable to the length of the central protruding conductor, in our case within the range of about 0.1-0.5 mm mm. We will show that this is not the case and a much better resolution can be achieved using a special signal difference method. This method is based on comparing results of two subsequent field measurements with the antenna displaced by a small distance Δl along its axis (Fig. 1 b). For planar microwave circuits the strength of the field is greatest close to the circuit surface and decays with increasing distance from the surface. For thin, short antenna ($d << l$, $l << \lambda$), placed in such a non-homogeneous field, the signal level can be approximated as a sum of the field contributions acting along the protruding conductor. If we split the antenna interaction areas into three regions - the antenna apex $\mathcal{A}$ middle section of the antenna protruding conductor $\mathcal{B}$ and the input to the shield C, the resulting signals I_1, I_2 before and after antenna displacement, can be formally written as

$$I_1 = I_1^A + I_1^B + I_1^C$$
$$I_2 = I_2^A + I_2^B + I_2^C \qquad (2)$$

We describe the signal levels in terms of the induced currents I_1, I_2 at the input to the coaxial shielding as the impedance of such short antennas are high in comparison with input impedance of the subsequent network and therefore the antenna functions more as a current source. Geometry and the position of the protruding conductor in the middle region $\mathcal{B}$ does not change with the antenna displacement and therefore the contribution from the same external field remains unchanged, $I_1^B = I_2^B$ For high-density planar structures, when condition (1) is fulfilled, the field

intensities rapidly decay with the distance increase above sample surface the field strength in the region C, and its contribution to the overall signal can be supposed to be negligible, $I_1^C = I_2^C = 0$. The measured signal difference

$$\Delta I = I_2 - I_1 = I_2^A - I_1^A \qquad (3)$$

depends only on the field solution and changes in the boundary conditions at the apex of the protruding conductor. As these changes are limited to the region Δl of the displaced antenna apex, the measured signal and the resolution of the measurement method are determined by the displacement Δl. In this way the field surrounding the conductor apex can be isolated and measured which improves the spatial resolution of the microwave field mapping.

The above analysis shows that one can improve the resolution by reducing the displacement Δl. At the same time this leads to a reduction in the signal level. Both mathematical simulations and the experimental results (see Fig 2) give a highly linear character of this dependency. The antenna is connected to the preamplifier and in a typical embodiment the antenna and the preamplifier form mechanically a single component. As a typical preamplifier is also highly linear, the measured voltage ΔU after its conditioning and transmission to the input of the acquisition instrument (VNA) is proportional to the antenna displacement Δl. We can therefore define the sensitivity of the system for a particular frequency by a single unit-less constant

$$S = \frac{\Delta I}{E_l \Delta l} \qquad (4)$$

Here E_l is the component of the electric field intensity of the microwave field parallel to the antenna axis. The sensitivity constant S can be determined from the calibration measurement in a well-defined field standard and can be used during the scanning process for the calculation of real values of the electric field intensity. Fig. 2 shows the dependency of the measured signal difference ΔU on antenna displacement Δl. The design of this specific antenna was optimised for higher sensitivity S at frequencies close to 4 GHz. The measurement was performed using a well-defined field standard.

The level of the acquired signal depends not only on the signal induced in the apex of the conductor but also on the efficiency of its matching to the input of the coaxial line, the properties of the preamplifier and the transmission of the signal to the acquisition system (usually a Vector Network Analyser (VNA)). As the signal level must exceed the noise level, the sensitivity of the antenna may effectively limit its resolution and make it dependent on minimal detectable field intensities. For small displacements Δl the apex of the protruding conductor functions as a near ideal current source and one of the main factors influencing the sensitivity is the matching of such a high-impedance source to the input of the coaxial line and subsequently to a preamplifier. To improve the matching efficiency, the input impedance of the coaxial line has to be increased using a quarter-wavelength transformer, a resonator or other impedance matching circuits.

Figs 3 and 4, a,b,c demonstrate the benefit of the invention. A PCB surface capacitor with a small separation gap between its fingers as shown in Fig. 3 has been fabricated to act as a demonstrator sample. The capacitor has been fabricated on a substrate with the dielectric constant $\varepsilon=10.2$ and a thickness $t=127$ microns. The width of the fingers is $w=40$ micron, they are separated by a gap $g=60$ micron. The terminal a of the capacitor is the input terminal connected to the microwave generator and the terminal b is the output terminal connected to the load. The protruding conductor

of the antenna has length 1mm and diameter 8μm. The antenna is driven using precision motorized stages very close to the sample surface with specified separation above the circuit. Figures 4 a,b represent scanned field images of the normal electric field acquired for two different antenna/sample separations of 6 μm and 12 μm respectively. Fig 4 c is the difference of these signals. We can clearly observe significant resolution enhancement for the signal difference. As the antenna is sensitive to the field acting along the entire length of the protruding conductor, the scattered field intensities at higher distances above the sample represent the main contribution to the level of the acquired signal. The signal difference corresponds to the local electric field intensities surrounding the antenna apex only. The signal difference also reveals weak local field intensities close to the signal lines of the bottom port of the capacitor, otherwise masked by strong background signals. These signals are induced by background fields acting along the whole length of the protruding conductor above the displaced apex of the antenna. A part of the result presented in Fig. 4 is also shown in Fig 5 in a more numerically readable format. This shows a line scan across the middle of the capacitor and considerable resolution enhancement achieved when the amplitudes of the signals 5 μm and 7 μm are subtracted from each other.

According to the invention one can also measure the phase of the electric or magnetic field. Indeed the signals presented in Fig. 4a and 4b are vectors and they are characterised not only by amplitude but also by phase. Therefore, their difference is also a vector that is in turn characterised by phase. This is demonstrated in Fig. 6 that correspond to the results presented in Fig. 5 respectively.

In the same way one can achieve enhancement of the resolution when measuring the amplitude and phase of the magnetic field. In this case one needs to use loop antenna instead of the coaxial probe. This is schematically shown in Fig. 7 for two positions of the loop antenna displaced with respect to each other along the vertical direction by Δl. Exactly for the same reason as in the case of the coaxial probe, it is advantageous to have the loop 3 of relatively long shape so that the shielding does not affect the signal detected by the loop and also does not affect performance of the DUT.

Fig. 8 demonstrate how the in plane (tangential) component of the electric and magnetic field can be measured. For these measurements the antenna is placed with an inclination of 45 degrees (for example) relative to the vertical axis. By rotating the antenna about the vertical (normal) axis, different spatial components can be measured. Standard Cartesian intensities, perpendicular and parallel to the surface of the DUT can be then recalculated. In the case of two measurements with the probe rotated by 180 degrees around the normal axis, a vertical and one tangential field intensity can be obtained.

$$E_z = \frac{1}{2\cos\alpha}(E_{0°} + E_{180°})$$
$$E_z = \frac{1}{2\sin\alpha}(E_{0°} - E_{180°})$$

(5)

Here $E_{0°}, E_{180°}$ are the electric field intensities before and after the rotation. Fig. 8 shows the position of the coaxial antenna before and after rotation by 180 degrees. After the rotation the antenna must be displaced along the horizontal direction so that the antenna's end is located above the same point of the DUT. This is shown in Fig. 8. For three measurements with the antenna rotated by 0, 120, 240 degrees all three components can be calculated.

$$E_x = \frac{1}{\sin\alpha}(2E_{0^\circ} - E_{120^\circ} - E_{240^\circ})$$

$$E_y = \frac{1}{\sqrt{3}\sin\alpha}(E_{120^\circ} - E_{240^\circ}) \qquad (6)$$

$$E_x = \frac{1}{3\cos\alpha}(E_{0^\circ} + E_{120^\circ} + E_{240^\circ})$$

In general the field can be elliptically polarised and the phase of the field intensities may vary for different spatial directions. Therefore, both the amplitude and the phase of the signal should be acquired by a phase-sensitive VNA and the intensities of the electric fields in (5) and (6) represent complex amplitudes of the signal. Fig. 8 shows electric field coaxial antenna. However, magnetic field antenna can be used for the measurements of the horizontal field components in the same manner.

The tangential components in Fig. 9 were acquired using a 45° inclined antenna. The measurements were performed above a microstrip line, the position of the strip edges are highlighted by dashed lines. The antenna was aligned in two opposite directions perpendicular to the strip. In this experiment the distance from the circuit surface was chosen to be relatively large (600 μm) as the tangential components are negligible close to the circuit surface and vanish at the conductive boundaries of the strip or ground plane. As the rotation about the vertical axis changes the probe's position above the sample surface, between the measurements the probe was offset so that the apex of the central protruding conductor is located at the same point for both measurements. For each direction two scans were performed with antenna displaced by 50 μm along the protruding conductor, their difference represent the field intensities at the antenna apex (Fig.9a). Normal E_z and transverse E_x electric field components, as presented in figure 9b were obtained using equation (5). As expected the transverse component E_x vanishes at the line center where the vector of electric intensity is perpendicular to the sample surface. This component has its maxima close to the strip edges with opposite directions of the field vector.

In one embodiment according to the invention, the means for the distance control are based on quartz crystal oscillator such as a tuning fork. Quartz crystals resonators are widely used in various electronic time measuring systems including electronic clocks and watches. They utilize piezoelectric properties of the quartz to convert electric signal to mechanical oscillations and vice versa. The strain of the vibrations causes the quartz to produce electric signal that results in changes in the effective impedance between two electrodes. The widespread use of the crystal resonators in the electronics is based on their high quality factor Q. When incorporated in the feedback of the oscillators, the extreme quality factor assures positive feedback with sufficient amplitude in a very narrow frequency range around their natural resonance frequency, resulting in an excellent stability for such a time base. To assure a high quality factor (typically above 10^4 of the crystal oscillators the air damping forces are reduced, and therefore the quartz resonators are encapsulated in a can and operate in low air pressure. In recent years quartz oscillators are more and more commonly used in other technologies completely unrelated to time measurements.

For example, a quartz tuning fork became commonly used in Atomic Force Microscopy (AFM) and Scanning Near-Field Optical Microscopy (SNOM), for control of the distance between the probe and the sample. The technique incorporates a dithered probe interacting with the surface in its proximity. The dependency of the amplitude and the phase of the probe's mechanical oscillation on

the probe/sample separation is used in a feedback to keep the separation constant. The tuning fork is utilized for the stabilization of the mechanical oscillations of the probe and the detection of the amplitude of the mechanical resonance. The method was originally introduced by Karrai and Grober [Karrai K., Grober R. D., Appl. Phys. Lett. 66, p. 1842 (1995)]; US patent 5641896 (Karrai). Various modifications of the system were proposed: with the probe oscillating either parallel or perpendicular to the surface [Tsai D. P., Yuan Y. L., Appl. Phys. Lett. 73 , p. 2724 (1998)], with both or only a single arm of the tuning fork [Kantor R., Lesnak M., Beredunov N., Shvets I. V., Appl. Surf. Sci 144-146 (1999), p. 510].

The oscillations in the tuning fork systems are usually excited by an external piezo-tube, bi-morph, thickness or shearing mode piezo plate, and not by the application of the signal directly on the fork electrodes. Thus the piezo-electric properties of the quartz tuning fork are disregarded. The reason for the use of an external dithering piezo is that the quartz resonator operates with a much lower quality factor (which drops by more than 2 orders of magnitude from their original value), caused by additional damping forces: air damping, non-elastic deformation within the system with the tip attached and drag forces of the tip/sample interaction. The ratio between the piezoelectric response signal and the amplitude of the excitation is directly proportional to the quality factor. Therefore, the level of the piezoelectric response with external excitation is also 2 orders of magnitude lower than that of standard quartz crystal applications and tend to be 10 - 100 times below the level of the amplitude of the excitation signal. Such low response is difficult to isolate from the original excitation signal, thus separate systems for mechanical dithering are usually used to electrically isolate both signals. Figure 10 represents the configuration with the dithering piezo and one vibrating arm oscillating parallel to the surface. The generator supplies an excitation signal to the thickness mode piezo. The generator also supplies a signal to the reference input of the LIA through phase shifter. The detection signal is collected from the electrodes of the tuning fork crystal and supplied to the input of the LIA. The topography tip is glued to the tuning fork crystal.

Self-excitation regime of the fork has been used by Chuang et al. [Chuang Y. H., Wang C. J., Huang J. Y., Pan C. L., Appl. Phys. Lett. 69 (1996), p. 3312] where a time-gating method was incorporated. In that method the excitation signal is multiplexed with an oscillation-sensing response by an electronic switch. This switch is triggered in sub-millisecond intervals, allowing separation of both signals. The disadvantage of this method, apart from the additional electronic instrumentation, is in the aliasing between the frequency of the oscillator and the frequency of the electronic switch gate. Conditions for the measurements have to be carefully chosen to avoid such aliasing, usually the frequency of the trigger signal has to be an order lower than mechanical resonance frequency. This results in a slower response and a longer time constant of the feedback system.

Our invention separates excitation and response signals. The separation is based on the observation that for a resonance frequency the mechanical oscillations (and electrical response) is shifted relative to the excitation forces (and the corresponding excitation electric signal) by 90 degrees that we shall call hereinafter the signal orthogonality. Those signals can be represented either by voltages on the fork electrodes or by currents flowing through the crystal. Due to relatively high impedance of the quartz crystal we will describe the signals by the currents: by the excitation current I_e and by the additional current I_f induced by mechanical vibrations of the tuning fork. The signal orthogonality allows phase-sensitive detection of the component corresponding to the mechanical vibrations only and suppressing the excitation signal. After the conversion of the currents to voltages this detection is performed using a Lock-In Amplifier (LIA) where the signals are demodulated relative to the reference coming from the excitation generator. The phase of either this reference or measured signals has to be adjusted in a phase shifter to assure 90 degrees phase

shift between the reference and the fork excitation signals. This phase shift allows nearly complete suppression of such an out-of-phase signal. The functional component I_f, corresponding to mechanical vibrations, is in-phase with the reference and it is fully demodulated. The amplitude of the output signal U_0 (for time $t = 0$) after lock-in detection can be written in the form

$$U_0 = \frac{R}{\tau} \int_{<<-\tau}^{t=0} [I_e \sin(\omega t) + I_f \cos(\omega t)] \cos(\omega t) e^{t''\tau} dt \cong \frac{R}{\sqrt{2}} I_f \quad (4)$$

where τ is the time constant of the lock-in demodulation and R is the conversion constant of the I/V converter. The output is proportional to the functional component I_f only and is used in the feedback to control tip/sample separation. The design of the probe does not require use of an external piezo element as both the excitation signal and the input for the response measurement are connected to the same electrodes of the tuning fork resonator.

According to the invention there are several alternative solutions to the issue of signal coupling to the quartz fork and phase-sensitive signal demodulation. In this disclosure we present two solutions by way of example. These solutions have been tested by us. As the correct choice of the signal phase is very important for maximising suppression of the excitation signal from the generator, incorporation of either a digital or analogue LIA with a high performance phase shifter is usually required. Any type of LIA (both analogue and digital) can be used, the signal from the generator is used as the reference.

Fig. 11 shows a simple circuit where the current induced in the tuning fork is measured. In this case the tuning fork is inserted between the generator and the IV converter. The voltage from the LIA is sent directly as a reference signal to the input of the LIA. The circuit itself functions as a phase shifter so that there is a 90 degree phase shift between signals coupled to the input and reference ports of the LIA.

For the excitation signal the circuit represents an electronic differentiator where the fork crystal functions as a capacitor. The voltage on the crystal electrodes is in phase with the voltage of the generator. The driving current is phase shifted by 90° and is suppressed by phase-sensitive detection of the lock-in amplifier. The functional component I_f, resulting from mechanical oscillations, is phase shifted relative to that of the excitation by an additional 90 degrees. Its resulting voltage $U_f = -RI_f$ has the phase opposite to that of the generator voltage and it is fully demodulated by the lock-in amplifier. Because the phase shift of the differentiator is constant for all frequencies, the source generator can be easily tuned to the resonance of the tuning fork system without any need for further phase adjustment. An additional advantage of this circuit is that the input of the I/V converter represents a virtual ground. It has virtually zero impedance relative to the ground point and brings very little uncertainties to the phase of the signals. Otherwise, the phase of the signals would be influenced even by small values of any resistive load as capacitance of the quartz fork is rather small (3-20 pF) and preamplifier with high input impedance hass to be applied, such as the one presented in patent EP 0864864 (Karrai). The operational amplifier is plugged very close to the quartz resonator and functions simultaneously as a signal conditioner whose low impedance output can be matched to the impedance of a standard transmission line. The suppression of the generator signal is typically more than 3 orders of magnitude, resulting in a residual signal of level 10-100 times smaller than that of the functional component. Such a ratio between the signals is sufficient for incorporation of the circuit in the feedback of tip/sample distance control system.

Fig. 12 shows another circuit of the quartz resonator plug-in where the voltage induced in the tuning fork is measured. The tuning fork is plugged between the output of the generator and the

input of LIA. In this case there is a separate phase shifter in the circuit to provide a 90 degree phase shift between the input and the reference ports of the LIA.

The distance control system based on the self-excitation of the tuning fork according to the invention has a number of advantages over the state-of-the-art system utilising the tuning fork and the dithering piezo. Both systems have similar sensitivities and comparable response times. However, the system based on the self-excitation has a simpler design, as no external dithering piezo is required. Also the system with the self-excitation is highly simple to adjust. There are no requirements of phase adjustment of the detected signal, as the electric excitation signal on the quartz fork electrodes is always in phase with the excitation forces and out of phase (shifted by 90°) with mechanical oscillations and response signal. In the state-of-the-art approach the phase shift, caused by the particular mechanical contact of the external dithering piezo and transfer of the mechanical vibrations from that piezo to the tuning fork is always present and the phase of the detected signal has to be tuned. These advantages result in an increase in the system's reliability and robustness.

When compared with the time gating technique as per [Chuang Y. H., Wang C. J., Huang J. Y., Pan C. L., Appl. Phys. Lett. 69, p. 3312 (1996)] the solution described in the present disclosure has also a number of advantages. It has a simpler electronic circuit, and does not require a time-gating modulation and complex adjustment of the frequency of the oscillator. According to the invention the circuit itself acts as a signal conditioner for the response signal thus resulting in a much better match in the input impedance. There is no residual modulation of the amplitude of the tip oscillation by the time-gating signal. These advantages result in a shorter response time and thus a faster scanning capability.

To maintain the separation between the antenna and the surface the following procedure is employed. First, the surface of the DUT is scanned using topography probe. The topography is recorded using e.g. shear force response and the self-oscillated tuning fork crystal as described above. The separation between the end of the probe and the surface of the DUT is then only in the range of 1-30 nm, which is the effective range of shear force. Then the topography probe is replaced for the antenna. We have found that advantageous way of doing this is as follows. Both topography probe and antenna are attached to a single XYZ translation stage. First, the topography probe is brought in focus of a long focal distance microscope. Then the computer controlled mechanical translation stage removes the topography probe from the focus and brings antenna to the same focal point of the microscope. The computer records the position offset as accomplished by the XYZ stage between the two positions. This offset is then added to bring the antenna in proximity of the surface. Additionally the Z-axis offset is added to increase the separation between the antenna and the surface from the value of 1-30 nm for the shear force-based topography probe to the value of e.g. 5 to 20 μm as required for the microwave signal acquisition.

The invention is used for scanning large size areas. Typically the size of the DUT could be in the range of some 50-200 mm. One can readily find precision motorised translation stages that are capable of proving accurate lateral displacement in this range along the X and Y axes. Indeed 0.1 micrometer precision for the lateral displacement is more than adequate. However, the situation with the height control, i.e. movement along the Z-axis normal to the DUT surface is much more complex. The problem is that one needs to have large dynamic range of Z-displacement, typically in the range of 10 mm or more and simultaneously, high resolution, down to some 1 nm. The high resolution is required as the probe height control is based on shear-force that is only active in the nanometer height range. Therefore, the conventional positioning tools used with the shear-force scanning microscopes are inadequate for the purpose of the Z-axis displacement. Typically piezo

tubes or piezo stacks are used. They are capable of providing the required resolution of the displacement but their dynamic range is limited to a fraction of a millimetre. According to the invention, we propose the use of the hybrid solution utilising both the piezo stack and the motorised translation stage according to the following routine. During the topography scan the motorised stage is always maintained at such a position that the piezo stack is kept in close to the middle of its dynamic range. For example, if the piezo stack is displaced from the middle of its dynamic range, by more than $+_-$ 25%, the motorized stage performs adjustment and moves the probe so that piezo stack is placed again in the middle of its range. In this way the feedback system can operate smoothly as piezo stack with fine position resolution and not the motorised stage is involved continuously in the z-position adjustment
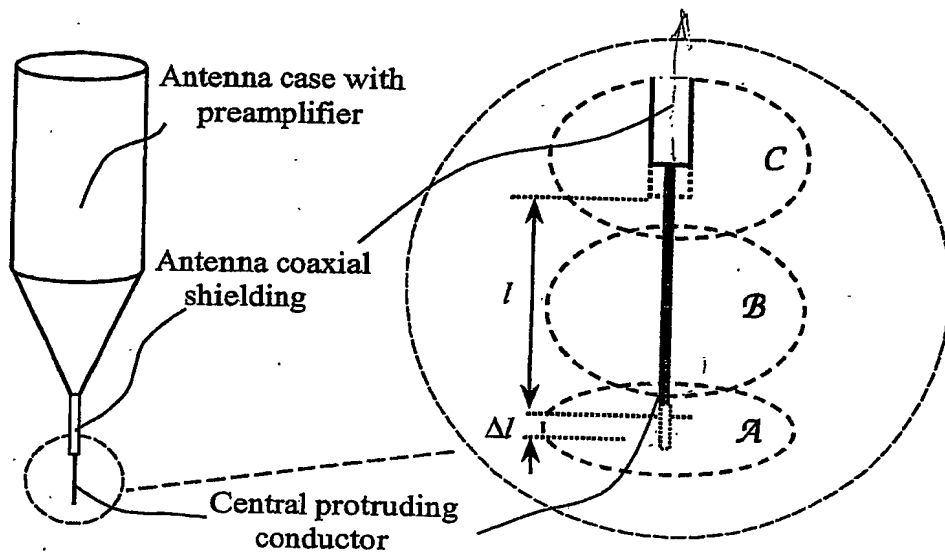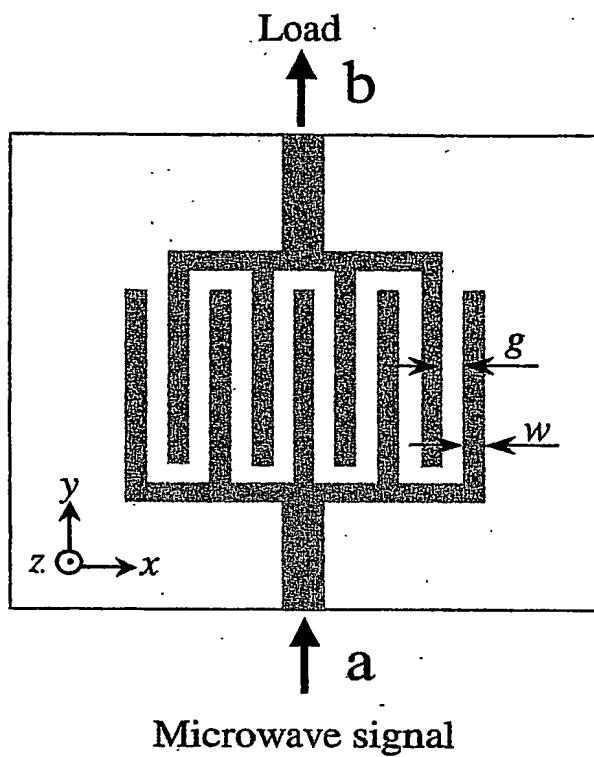
Antenna case with preamplifier

Antenna coaxial shielding

Central protruding conductor

C

B

A

l

Δl

Fig. 1a

Fig. 1b

Fig. 3

Fig. 4a   Fig. 4b   Fig. 4c

Fig. 5

Fig. 6

shielding

$B_C = 0$

loop

$B_{\mathcal{B}}$

Δl

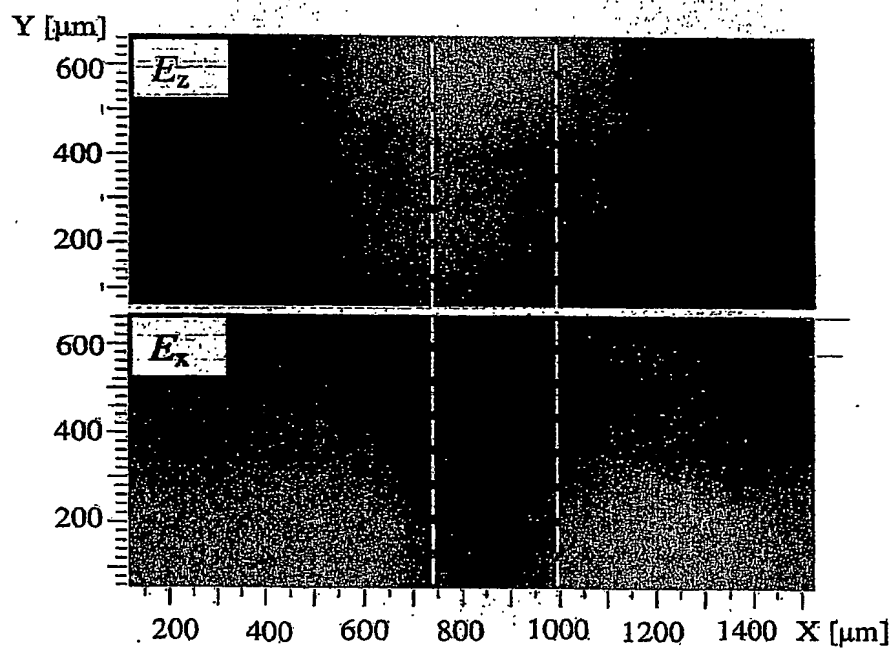effective area
of the antenna

$B_{\mathcal{A}}\ \otimes$
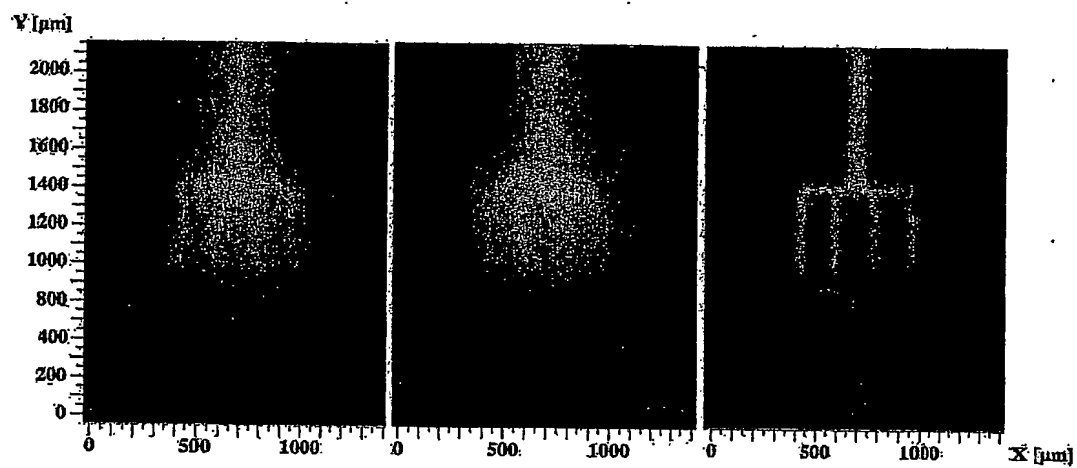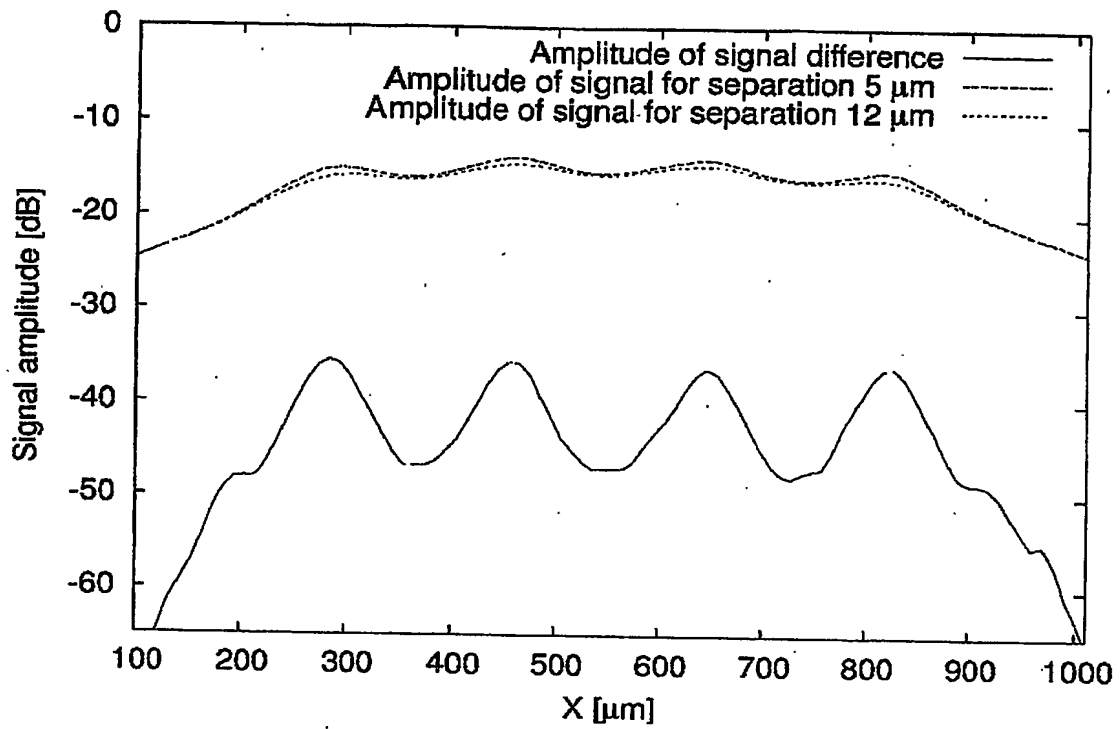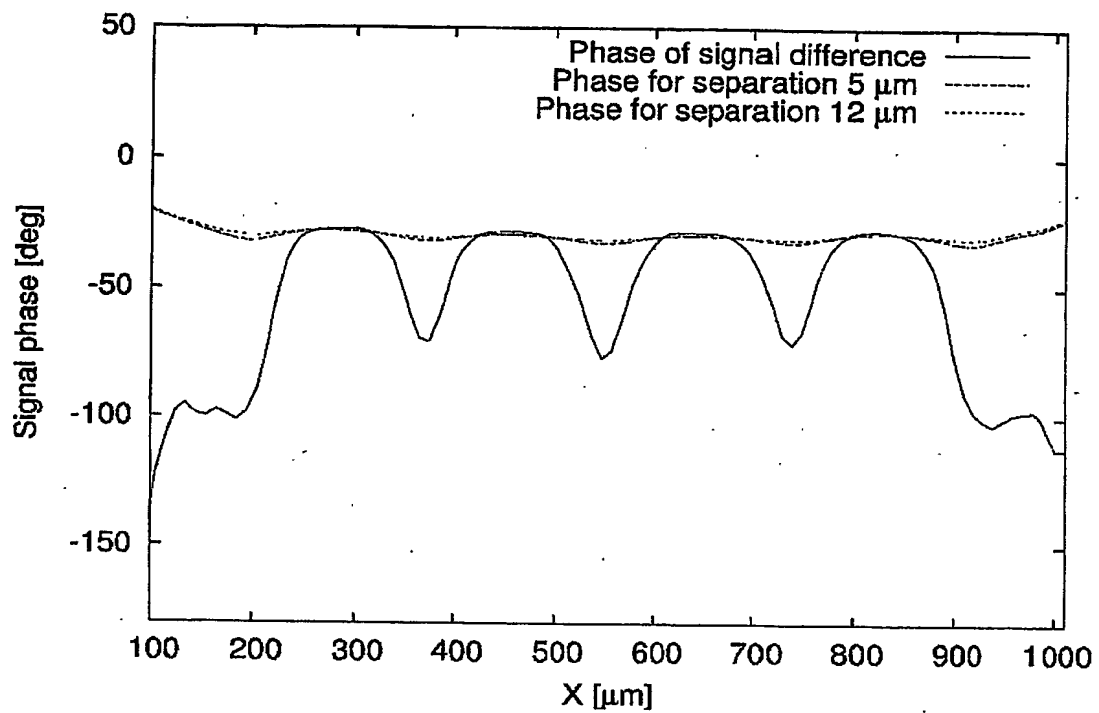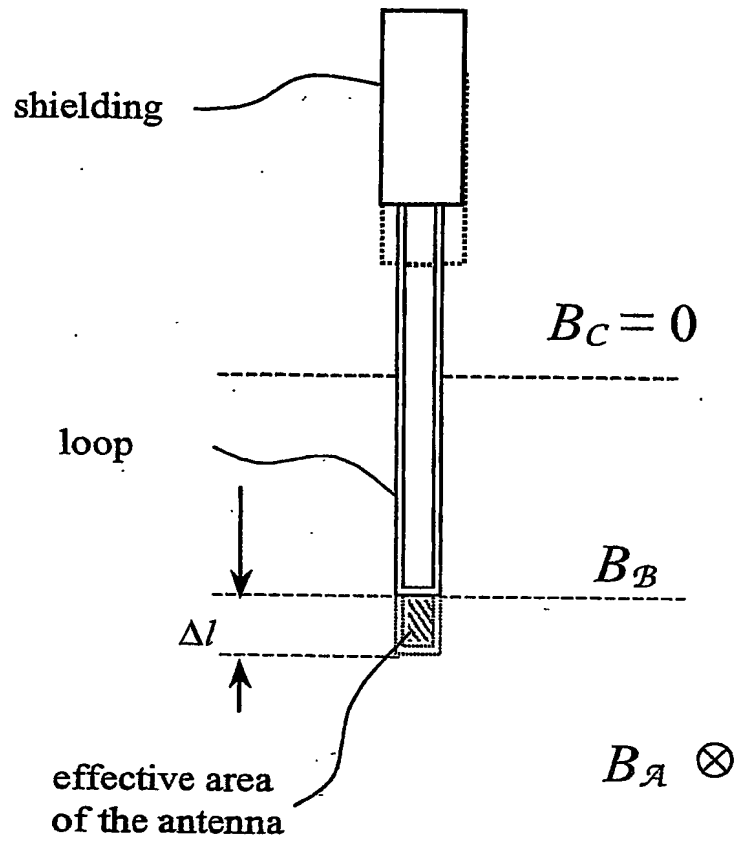
Fig. 7
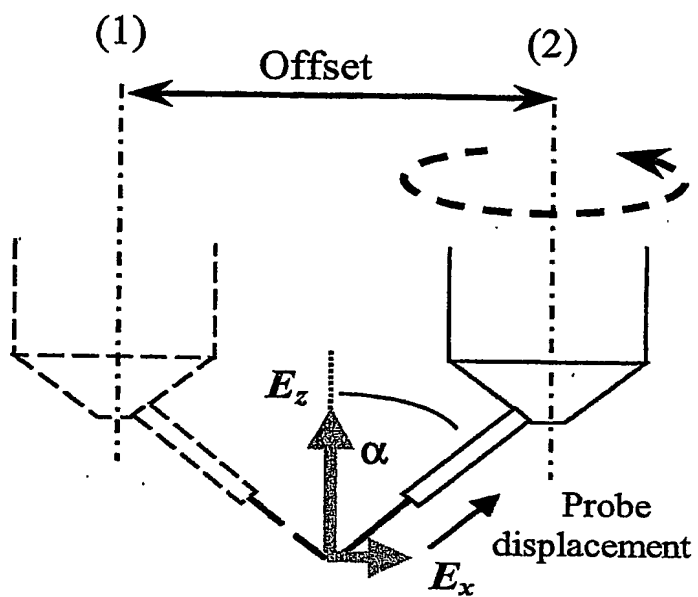
Fig. 8
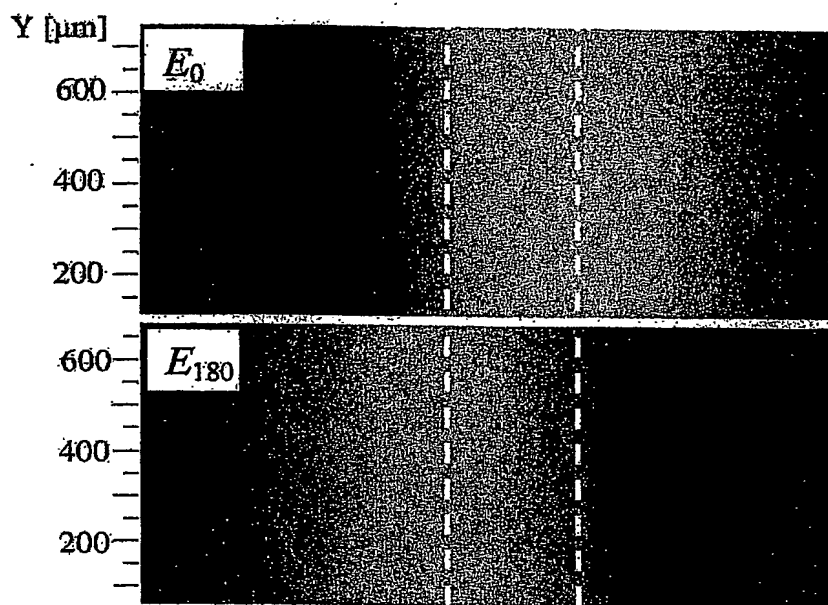
Fig. 9a
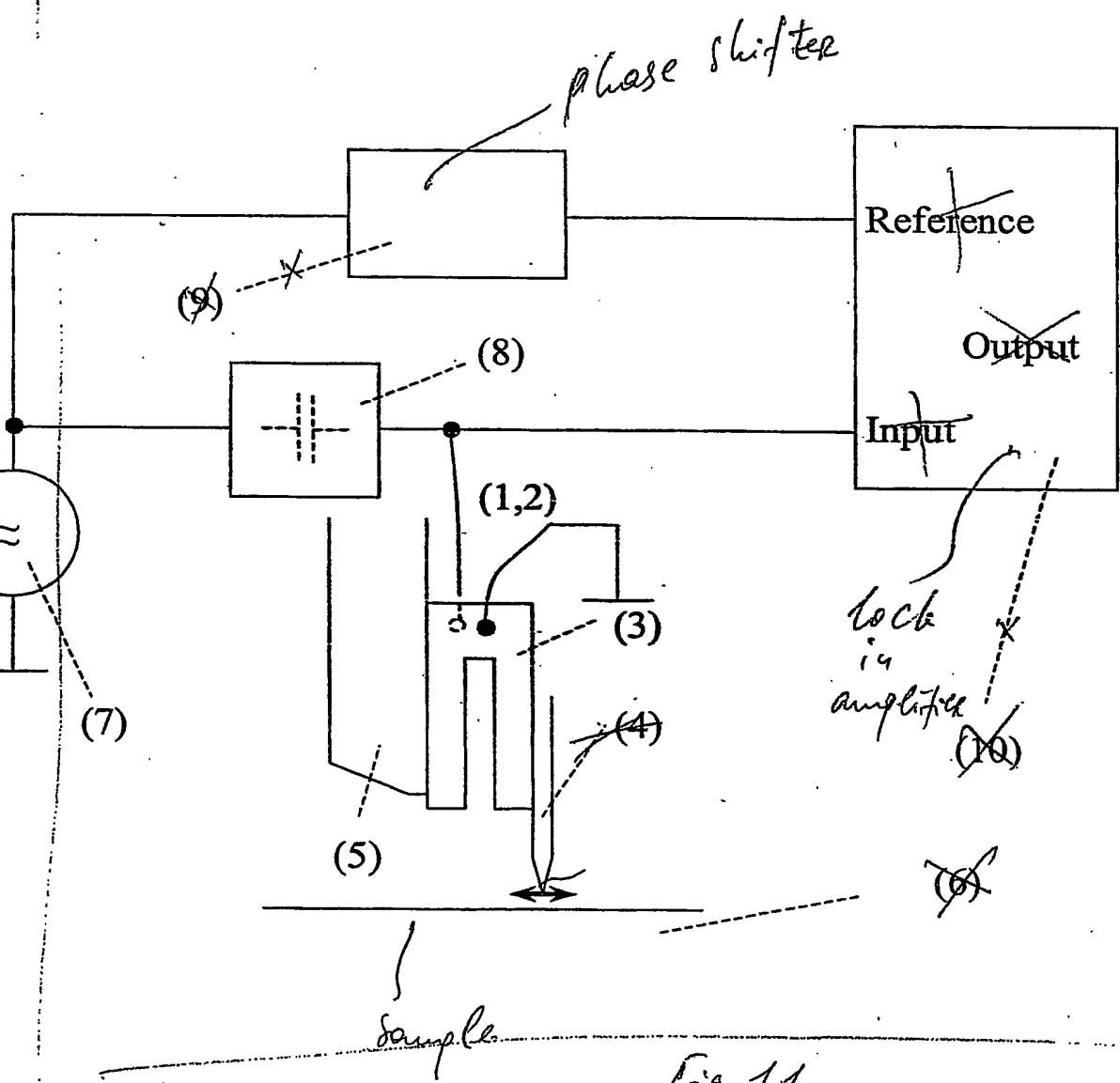
Fig. 9b

(1) Excitation signal
(2) Detection signal of tuning fork dithering
(3) Crystal: a tuning fork resonator
(4) Topography tip
(5) Tuning fork holder
(3) Excitation piezo
(7) Sample
(8) Generator – excitation source
(9) Phase shifter
(10) Lock-in amplifier
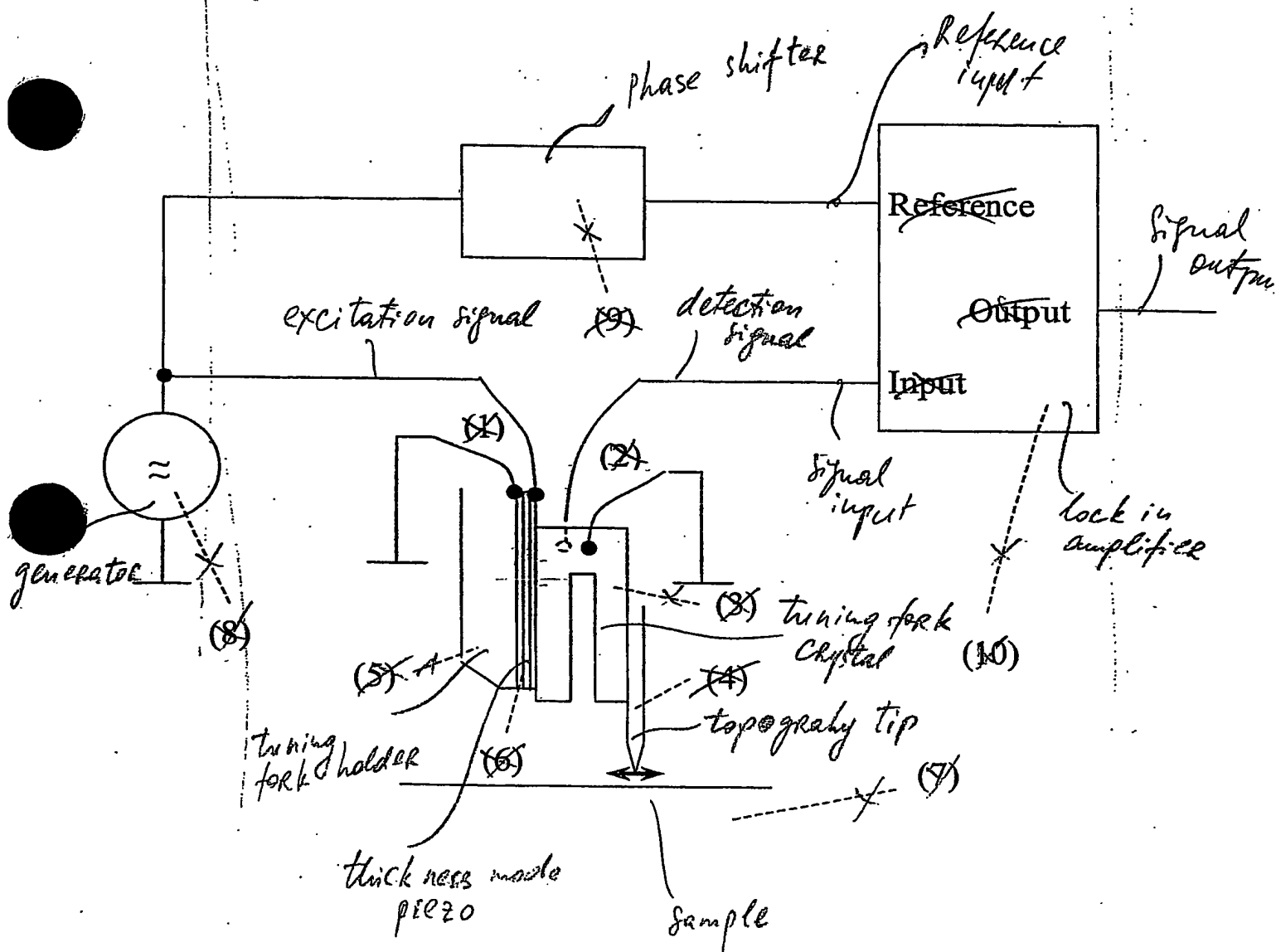
phase shifter

Reference input

Reference

excitation signal

(9)

detection signal

Output

signal output

Input

(1)

(2)

signal input

lock in amplifier

generator

(8)

(8) tuning fork Crystal

(10)

(5) A

(4)

topograhy tip

tuning fork holder

(6)

(7)

thickness mode piezo

sample

Fig 10

(1,2) Excitation and detection signal of tuning fork dithering
(3) Crystal: a tuning fork resonator
(4) Topography tip
(5) Tuning fork holder
(6) Sample
(7) Generator — excitation source
(8) Signal coupler
(9) Phase shifter
(10) Lock-in amplifier

Fig. 11.

(1,2) Excitation and detection signal of tuning fork dithering
(3) Crystal: a tuning fork resonator
(4) Topography tip
(5) Tuning fork holder
(6) Sample
(7) Generator — excitation source
(8) Operational amplifier plugged as I/V converter
(9) Resistor for I/V conversion
(10) Lock-in amplifier

Fig. 12